N88-25723

PERFORMANCE OF THE ICAO STANDARD "CORE SERVICE" MODULATION AND CODING TECHNIQUES

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ABSTRACT

There is significant interest in the aviation community in establishing a satellite communications system architecture to provide comprehensive aeronautical communications services. Incorporated into the architecture is a "core service" capability, which provides only low rate data communications (e.g., 300, 600, 1200 information bps), to allow for basic air traffic services and aeronautical operational control. In November 1986, the International Civil Aviation Organization's (ICAO) Special Committee on Future Air Navigation Systems agreed upon a standard modulation technique for the core service. The new standard, designated aviation binary phase shift keying (A-BPSK), is essentially differentially-encoded symmetric binary phase shift keying with square-root 40 percent raised cosine pulse shaping. It was also agreed that the recommended coding technique is an interleaved constraint-length 7 convolutional code with rates 1/2 or 3/4. This paper contains material that was given in working papers to ICAO and to the Airlines Electronic Engineering Committee (AEEC). In addition, it contains more recently generated material dealing with the application of maximum likelihood principles to the problem of detection of A-BPSK modulated data, that was transmitted over a Rician fading channel.

AVIATION BINARY PHASE SHIFT KEYING

A-BPSK is a form of differentially-encoded filtered symmetric BPSK (Schwartz et al., 1966). Symmetric BPSK differs from conventional BPSK, in that every second symbol is transmitted in the quadrature channel. Note that this scheme is still binary phase shift keying in the sense that the phase, of any given transmitted symbol, can take on only one of two possible values. Therefore, symmetric BPSK exhibits the same robustness to phase noise and nondispersive fading as conventional BPSK. However, while conventional BPSK has phase transitions of π radians, symmetric BPSK has phase transitions of $\pi/2$ radians. Therefore, symmetric BPSK exhibits less spectral regrowth after filtering and hardlimiting, than does conventional BPSK.

Pulse shaping, using members of the family of square-root raised cosine filters, was studied by computer simulation. It was found that, for a wide range of roll-off factors, the resulting transmitted signal has relatively small envelope variations. The root-mean-square and extreme values of these variations were computed as a function of roll-off factor. The envelope variations are smallest in the vicinity of 40 to 50 percent. Therefore, it seems reasonable to expect that A-BPSK, which has a roll-off factor of 40 percent, will be robust to nonlinear amplification. The power spectral density of A-BPSK, after hardlimiting, can be seen in Figure 1. While a considerable number of sidelobes have appeared, the spectrum is still much more compact than that for conventional BPSK. In fact, the spectrum is only marginally worse than that for MSK!

The performance of coherently-detected A-BPSK has been evaluated using computer simulation, for a number of different scenarios. For a variety of different fading bandwidths and K-factors, the performance of hardlimited A-BPSK is shown in Figure 2. Here, the K-factor is defined to be the ratio of the average power of the Rayleigh fading signal path to the power of the direct path. For these simulations, the symbol timing was assumed to be perfect and the carrier recovery circuit was assumed to be locked to the direct path signal. Within the accuracy of the simulations, the results in Figure 2 are practically identical to the corresponding ones that were generated for differentially-encoded coherently-detected BPSK (DECPSK).

Performance has also been evaluated for the rate 1/2 coded case. The coded performance for both static channel conditions and Rician fading conditions, with a K-factor of -7 dB, is shown in Figure 3. For these simulations, the interleaving depth is sufficiently large that there is negligible interleaving loss. The magnitude of the "soft decision" was taken to be the smallest of the two sample magnitudes affecting the given decision. Note that the presence of fading degrades the performance by less than 1 dB, over the range shown.

A PARTIALLY COHERENT APPROACH TO THE DETECTION OF A-BPSK

As was shown in the previous section, A-BPSK performs very well even when it is hardlimited. Hardlimited A-BPSK can be thought of as a type of continuous phase modulated (CPM) signal. Fortunately, attractive discrete-time receiver structures, based on maximum likelihood principles applied to Rayleigh fading channel transmission, exist for the reception of CPM signals. In this section, the application of such a receiver structure to the reception of A-BPSK is discussed.

Consider the case of Rayleigh fading transmission where the receiver down-converts the CPM signal to complex baseband, passes it through an ideal anti-aliasing filter, and then samples it. The filter bandwidth, B, is wide enough that for practical purposes all of the signal energy is passed, including the signal energy that is spread by the fading process. Furthermore, it is assumed that the filter's transfer function possesses the appropriate symmetries so that the additive Gaussian noise is still white after the sampling is performed. Once these assumptions are accepted, it is sufficient to perform all of the modelling and analysis using the discrete-time complex baseband model.

Consider the case of maximum likelihood sequence estimation, where each of M possible data sequences corresponds to transmitting a distinct constant envelope signal, with each distinct signal having a duration of N samples. Furthermore, let the amplitude of the signals be normalized so that the magnitude of each complex sample is one. Therefore, the set of possible transmitted signals can be denoted as the set of N-vectors,

$$X = \{x_1, \; \ldots \; , x_M\}. \tag{1}$$

Let the corresponding signal at the output of the discrete-time channel be denoted by the N-vector $\mathbf{y}$. Then the maximum likelihood detection choice is

x_k, if

$$p(y|x_k) > p(y|x_m) \qquad ; m = 1, \ldots, M ; m \neq k. \tag{2}$$

The detector must determine the M conditional probabilities, or M equivalent metrics. Consider the case, where the first operation performed by the detector is to compute M new vectors given by

$$y'_m = P_m y \qquad ; m = 1, \ldots, M, \tag{3}$$

where P_m is a diagonal matrix with ii'th element being the i'th element of x^*_m, and $*$ denotes the complex conjugate. This operation can be viewed as removing the assumed modulation from the received signal, for each of the M hypotheses. It is easy to show that equations (3) represent a set of linear transformations, each with a Jacobian of magnitude one. Therefore,

$$p(y'_m|x_m) = p(y|x_m). \tag{4}$$

Let R be the Toeplitz covariance matrix with its elements being the normalized autocorrelation coefficients that correspond to the power spectral density function, $S'(f)$, of the received signal when the input to the channel is an unmodulated carrier. It can be shown that choosing the hypothesis with the maximum likelihood is equivalent to choosing the hypothesis that minimizes the quadratic metric given by,

$$J_m = y'^H_m R^{-1} y'_m \qquad ; m = 1, \ldots, M. \tag{5}$$

where superscript H denotes the complex conjugate transpose.

Next, a theorem will be given that forms the basis for the detection strategy. It is assumed here that $S'(f)$ is symmetrical about the origin so that the elements of R are real. Extension to the case where the elements of R are complex is straightforward but tedious.

THEOREM: Given the normalized autocorrelation coefficients $1, r_1, \ldots, r_{N-1}$, let R be the positive definite Toeplitz matrix with ij'th element $r_{|i-j|}$ $(r_0 = 1)$. Then

$$R^{-1} = A^T D A, \tag{6}$$

where D is a diagonal matrix with ii'th element $(V_{i-1})^{-1}$, and A is the forward prediction matrix given by

$$A = \begin{bmatrix} 1 & 0 & 0 & \ldots & 0 \\ a_1^1 & 1 & 0 & \ldots & 0 \\ a_2^2 & a_2^1 & 1 & \ldots & 0 \\ \vdots & & & \ddots & \vdots \\ a_{N-1}^{N-1} & a_{N-1}^{N-2} & a_{N-1}^{N-3} & \ldots & 1 \end{bmatrix}. \tag{7}$$

Here, a_j^i is the i'th coefficient of the j'th order linear predictor, and V_j is the corresponding normalized prediction error, with V_0 defined to be one.

From the computational point of view, well known recursive algorithms exist that allow one to compute the predictor coefficients and the expected normalized prediction errors, from either autocorrelation data or coefficients of the all-pole filter with the spectrum S'(f) (Makhoul, 1975).

This theorem has an intuitively pleasing interpretation. It implies that the maximum likelihood hypothesis is the one that minimizes the sum of the weighted squared forward prediction errors, where each of these prediction error terms is weighted by the inverse of its expected value. Clearly, the prediction filters have orthogonalized the metric given in equation (5).

An interesting case is the one where S'(f) can be closely approximated by an L'th order all-pole filter, with L much smaller than N. Note that for an L'th order model, all linear predictors of order L and greater have the same coefficients and the same expected normalized squared prediction error. Therefore, except for the first L samples, the forward error prediction filter is a time-invariant finite impulse response filter of order L. Furthermore, assuming that the modulator has a finite memory, each squared term (i.e. prior to the summation) is dependent only upon a finite number of adjacent bits. This is exactly the situation that can be efficiently handled using dynamic programming. Often, the combined memory of the modulator and channel can result in a prohibitively large number of states. In this case, a decision feedback, truncated state, or reduced state approach can be adopted.

In theory, the above receiver concept can easily be extended to include Rician fading. However, in practice, the receiver would have to accurately estimate the gain and phase of the direct path, which is undesirable. Here, a different approach is taken. The direct path, in the assumed channel model, is replaced by a slow Rayleigh fading path. The assumed S'(f) for the computation of the receiver parameters is shown in Figure 4. In the simulation, the fading spectrum is Gaussian, whereas it is approximately rectangular in the assumed spectrum. Thus, there are several mismatches between the simulated channel and the assumed one, that result in a departure from true maximum likelihood detection. Further departure occurs when steps are taken to reduce the receiver complexity. The assumed composite spectrum is approximated by an eighth-order all-pole filter, with the resulting spectral characteristic shown in Figure 5. Also, the number of states was reduced to 4 by using decision-feedback detection. The resulting performance of the partially coherent receiver is illustrated Figure 6. For bit error rates less than 10^{-2}, the partially coherent receiver noticeably outperforms the ideal DECPSK receiver. However, for the higher bit error rates, typical of operation with convolutional coding, there is little difference between the performance of the two schemes. Unlike the case for DECPSK, the partially coherent receiver does not require carrier phase recovery, and requires only approximate carrier frequency estimation.

CONCLUSIONS

A-BPSK was described and simulated performance results were given that demonstrate robust performance in the presence of hardlimiting amplifiers. The performance of coherently-detected A-BPSK with rate 1/2 convolutional coding was given. The performance loss due to the Rician fading was shown to be less than 1 dB over the simulated range. A partially coherent detection scheme, that does not require carrier phase recovery, was described. This scheme exhibits similar performance to coherent detection, at high bit error rates, while it is superior at lower bit error rates.

REFERENCES

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Makhoul, J. 1975. Linear Prediction: A Tutorial Review. Proc. IEEE, vol. 63, pp. 561-580.

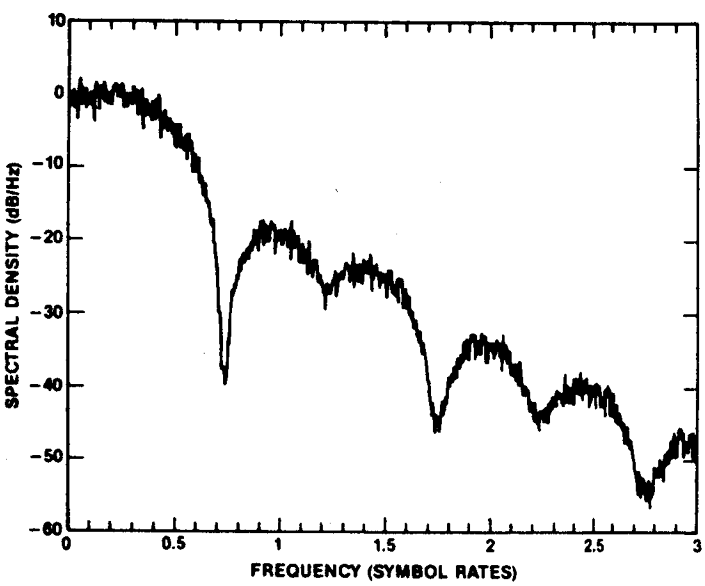

Fig. 1. The power spectral density of hardlimited A-BPSK.

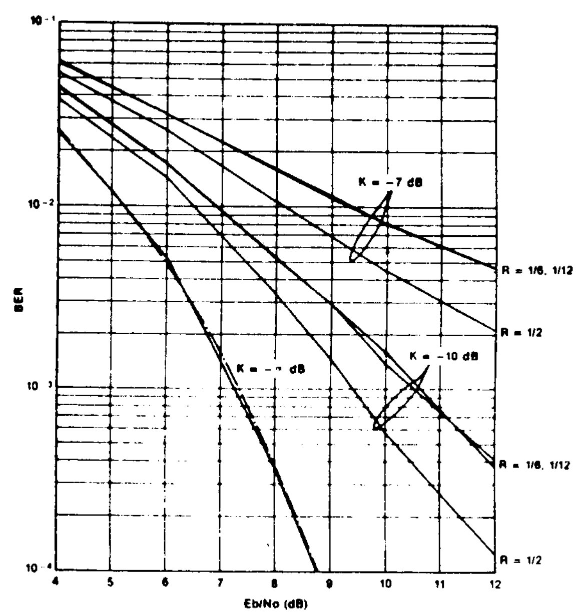

Fig. 2. The performance of hardlimited A-BPSK transmitted over a Rician fading channel. The fading rate, R, is expressed as a fraction of the bit rate.

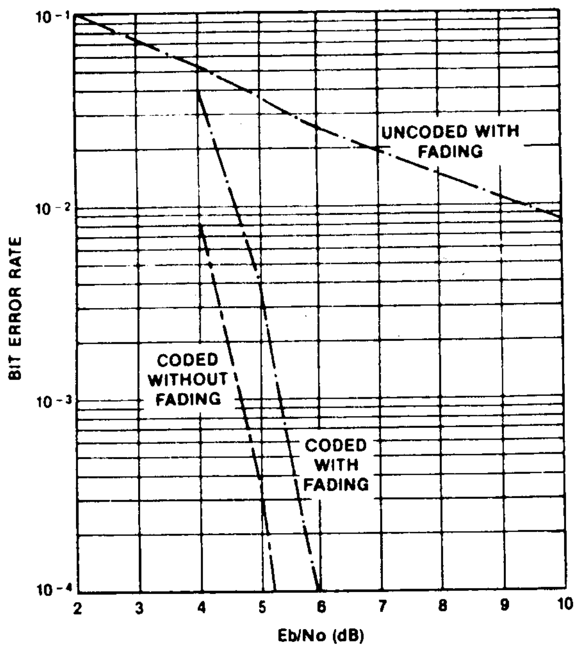

Fig. 3. The performance of A-BPSK with rate 1/2 coding, transmitted over static and Rician fading (K= -7 dB) channels. Here, the fading rate is 1/12 of the bit rate.

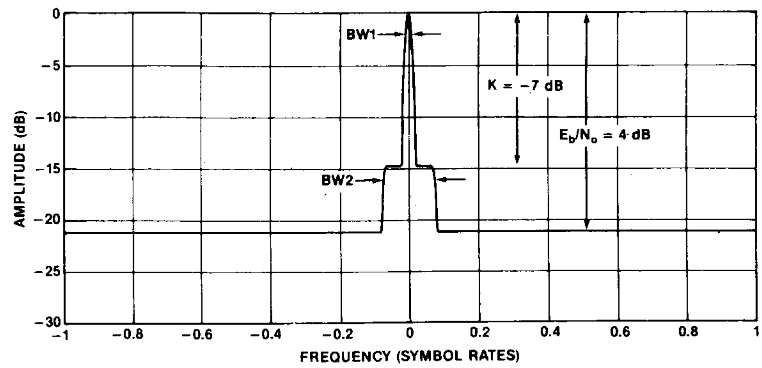

Fig. 4. The assumed composite spectrum, S'(f).

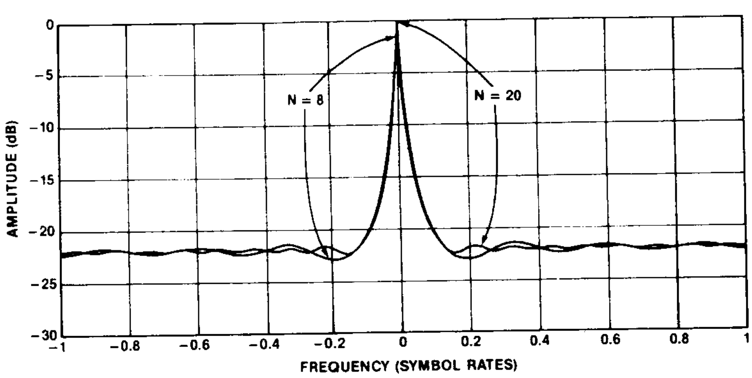

Fig. 5. The 8'th-order and 20'th-order approximations of the assumed composite spectrum.

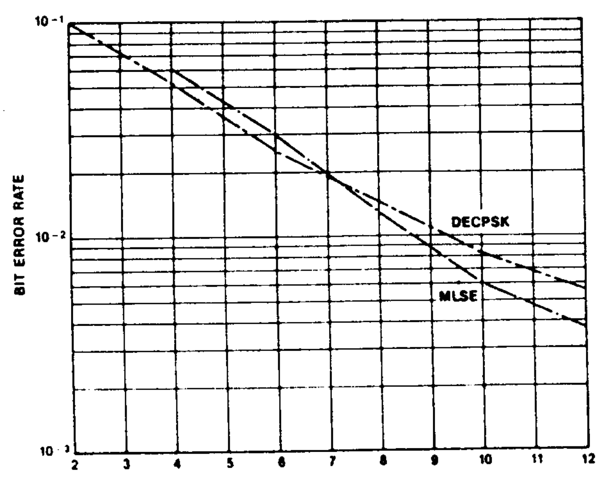

Fig. 6. A comparison of the performance of DECPSK and the partially coherent detection scheme, transmitted over a Rician fading channel (K = -7 dB).